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Spallation Neutron Source Beam Current Monitor Electronics[1]

M. Kesselman, W. C. Dawson

Brookhaven National Laboratory

Abstract. This paper will discuss the present electronics design for the beam current monitor system to be used throughout the Spallation Neutron Source (SNS) under construction at Oak Ridge National Laboratory. The beam is composed of a micro-pulse structure due to the 402.5MHz RF, and is chopped into mini-pulses of 645ns duration with a 300ns gap, providing a macro-pulse of 1060 mini-pulses repeating at a 60Hz rate. Ring beam current will vary from about 15ma peak during studies, to about 50Amps peak (design to 100 amps). A digital approach to droop compensation has been implemented and initial test results presented.

BACKGROUND

This is the fourth[1,2,3] paper presented on this subject. The earlier papers describe the decision process to attempt to use identical electronics throughout to provide a compatible approach to system diagnostics. A PC based instrument design philosophy was adopted for the diagnostics wherever it would apply, and a compromise in transformer frequency response, droop, and standardization resulted in a decision to use Bergoz® FCT (Fast Current) transformers. Individual areas within the SNS will have current transformers that suit the dimensional requirements, while maintaining electrical performance compatible with required performance and the system electronics. All electronics would be identical and could be placed anywhere in the system. Therefore, the electronics has been designed with flexibility in its configuration.

CHALLENGES

There were a number of major challenges to be addressed:

- Measurement of chopper characteristics vs. Ring turn by turn current.
- Goal of a single design to minimize design cost, and provide interchangeability
- Large dynamic range in Ring and RTBT
- Baseline restoration and droop compensation
- Integration accuracy vs. sampling rate
- Noise, response characteristics, filtering and digitizing aliasing
- Testing

[1] Work performed under the auspices of the U.S. Department of Energy

•Calibration

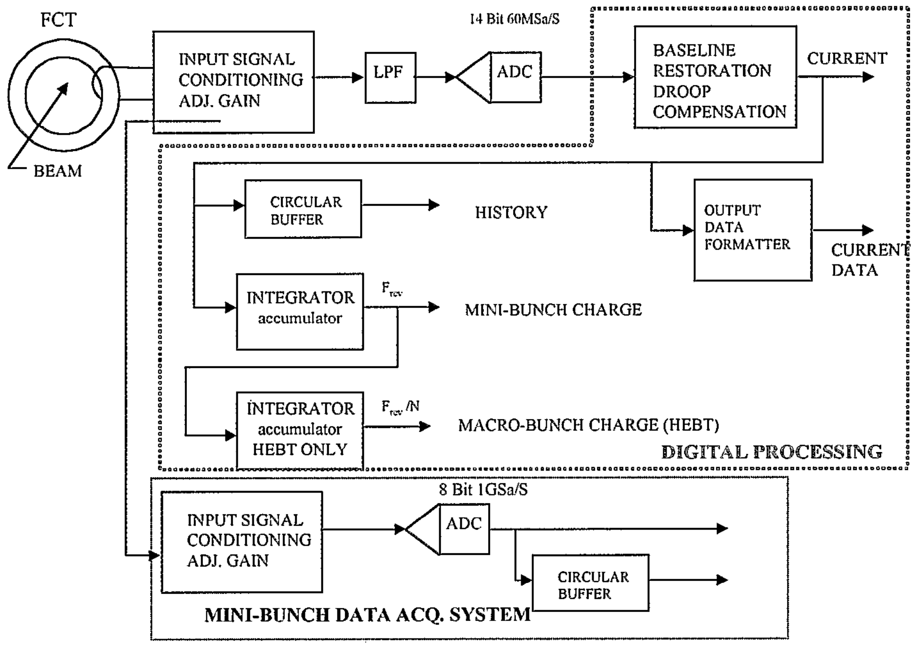


FIGURE 1. Block diagram of BCM electronics. This shows two digitizing systems. The faster digitizer is a separate system used for analysis of chopper characteristics and is intended for short time samples. The slower speed digitizer is intended for general current monitoring and has a reduced bandwidth. Baseline restoration and droop compensation are accomplished digitally in the slower system.

DETAILS

The selection of the FCT transformer permits high speed response, while allowing droop to be compensated digitally. The electronics will provide a broad-band output suitable for observing the chopper characteristics. Provisions for jumper selected configurations for the lower current MEBT, Linac, and HEBT and the higher current Ring and RTBT allow the same electronics to be used throughout the SNS. The large dynamic range in the Ring and RTBT is handled by employing different gain paths that are selected by switchable amplifiers. These amplifiers (OPA680 series) switch in less than 100ns, allowing them to be switched during the "gap" time in the Ring. The large voltages expected (25 to 50 volts) are sufficient to cause sensitive amplifiers to fail. Therefore, the use of protected amplifiers is necessary in the Ring and RTBT. A system block diagram is shown in figure 1. The input signal conditioning circuitry will separate the signal into two paths. The broadband path will be sent to a separate high-speed digitizer capable of analyzing the chopper characteristics. The second path will be channeled to a signal processing system capable of amplifying the signal adequately for digitization by a 14 bit 68MSa/s digitizer. This digitizer is synchronized with the revolution frequency of the Ring ($64xF_{rev}$).

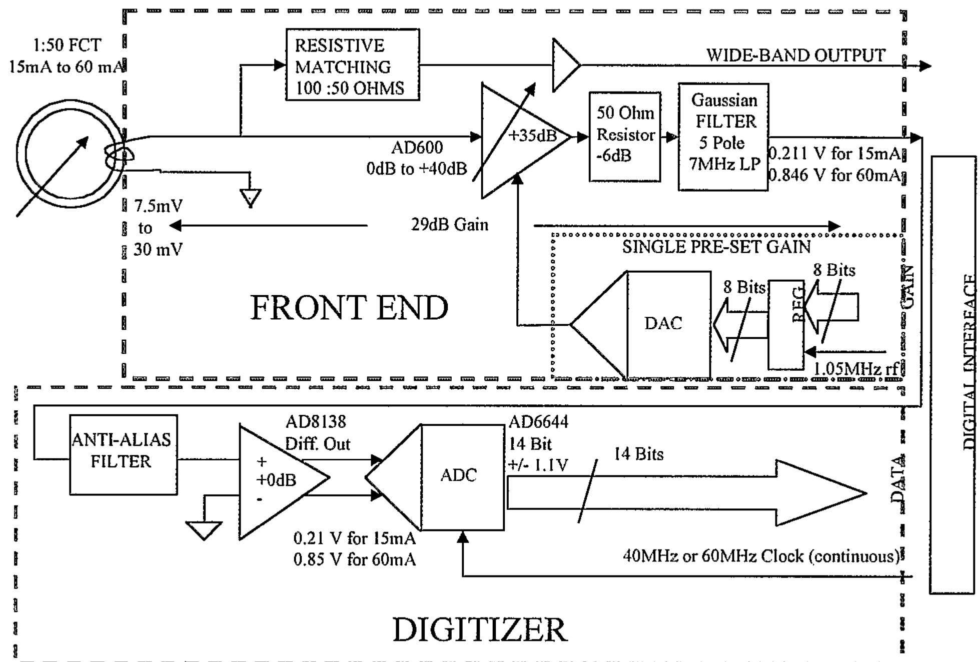


FIGURE 2. Block diagram of BCM electronics used for the MEBT, Linac and HEBT. A 100Mhz bandwidth output is provided for monitoring purposes. Jumper selection configures this processing configuration.

MEBT, Linac, and HEBT Analog System

A high gain configuration is jumper selected to accommodate the lower currents expected in the Front End, MEBT, Linac and HEBT, and is shown in figure 2. A variable gain controlled amplifier (AD600) is used to amplify signals in the 0 to 25mV range (for 0 to 50ma) to a voltage range of 0 to 500mV for the ADC. A factor of two in headroom has been reserved for current peaking. The wideband transformer signal is buffered by an amplifier with bandwidth >100MHz (OPA3680).

The Ring rise time is expected to deteriorate to about 50ns. This requires a 7MHz bandwidth. Increasing the bandwidth increases the noise and the resolution degrades. Therefore, a 5 pole - 7MHz Gaussian filter was chosen to minimize overshoot, and provide significant filtering for aliasing considerations. This filter will provide about 42dB attenuation at 34MHz. Additional attenuation of 34dB by a 5 pole – 17MHz, 0.01dB, Chebyshev filter and two additional amplifier stages limited to 34MHz provides >80dB attenuation at the Nyquist frequency.

Ring and RTBT Analog System

To accommodate the large dynamic range of the Ring and RTBT, the configuration of figure 3 is employed by a jumper selection. This configuration employs protected high gain stages and a number of lower gain paths that are digitally selected to establish a variable gain capability. The paths are switched by fast switching amplifiers. Switching during "gap" time will allow for no loss of turns.

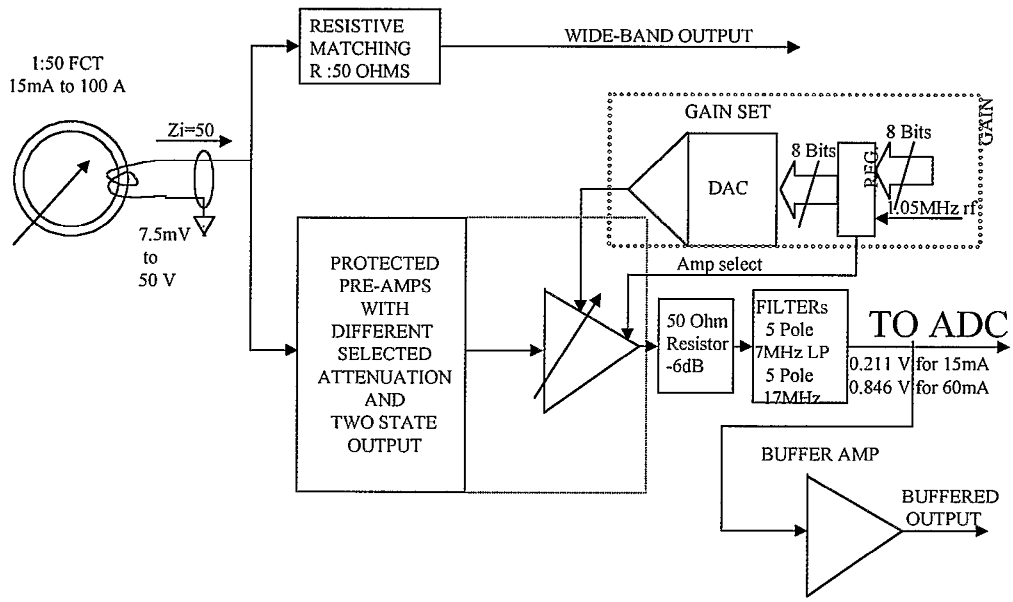
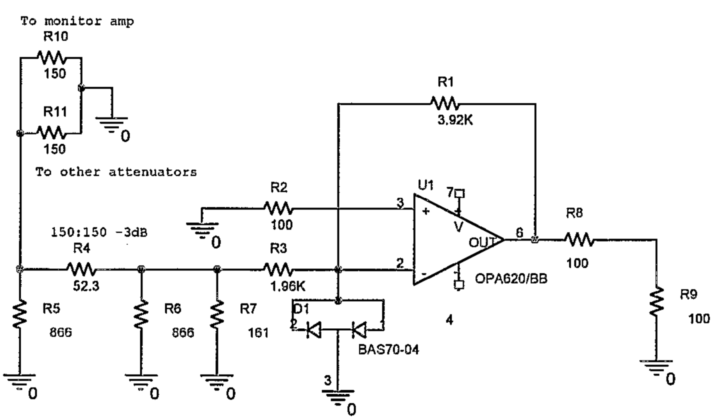


FIGURE 3. This shows the general block diagram for the jumper selected configuration for the Ring and RTBT BCM electronics. A protected amplifier system handles the high voltage levels expected in the Ring and RTBT. Different amplifier gain paths are selected digitally to permit handling the 1000:1 dynamic range.



For the attenuator:

$\Delta Z_i/ Z_i=2*(\Delta Z_L/ Z_L)/[2*N+(N-1)*(\Delta Z_L/ Z_L)]$; N= power ratio=$P_i/P_o$

$Z_L=161||Z_i$; $Z_i = V/I = V/(V-V_d)/1.96K \sim 1.96K*(1+ V_d/V)$

For $V_d/V=1/50=0.02$; $\Delta Z_L=149.000648-148.7788 = .2218$ and $(\Delta Z_L/ Z_L)=0.001491$

For the 3dB pad N=2 and $\Delta Z_i/ Z_i= .074\%$

FIGURE 4. A protected amplifier must both protect the amplifier against high input signal voltage as well as assure that input impedance is maintained constant.

Protected Amplifier

Signal levels in the Ring can get as high as 25 Volts for 50 amps. If one considers doubling this for headroom, the signals are clearly too large for the amplifiers to handle. Therefore, the high gain paths must be protected. The amplifier shown in figure 4 provides both protection and assures a near constant input impedance. The input impedance must not vary by more than 0.1% to assure proper accuracy. This is achieved by providing an input resistance of more than 1K Ohm before the protection diodes. In so doing the diode distortion for high amplitude signals is not reflected to the other gain paths, however, noise is increased. One of the 150 Ohm input resistors represents a 150 Ohm attenuator chain. The attenuation is selected such that after the first stage of attenuation the signal is maintained below +/- 2.0 V (max input signal for the AD600).

Adjustable Gain Stage

The gain changing configuration is shown in figure 5. The attenuation provided is also shown. The amplifiers are part of a triple op-amp (OPA3680). This amplifier is a voltage feedback type amplifier with a digital control input permitting it to switch "off" within 100ns. The amplifiers of figure 5 are all of this type. The switched amplifier is configured as a +2 gain amplifier, providing no signal punch through when switched "off". The summer, adds the outputs of each path (all paths "off" except one). The system resolution has been established at 0.5% of full scale, allowing 4 gains. A table of gains and expected resolution is shown in Table 1.

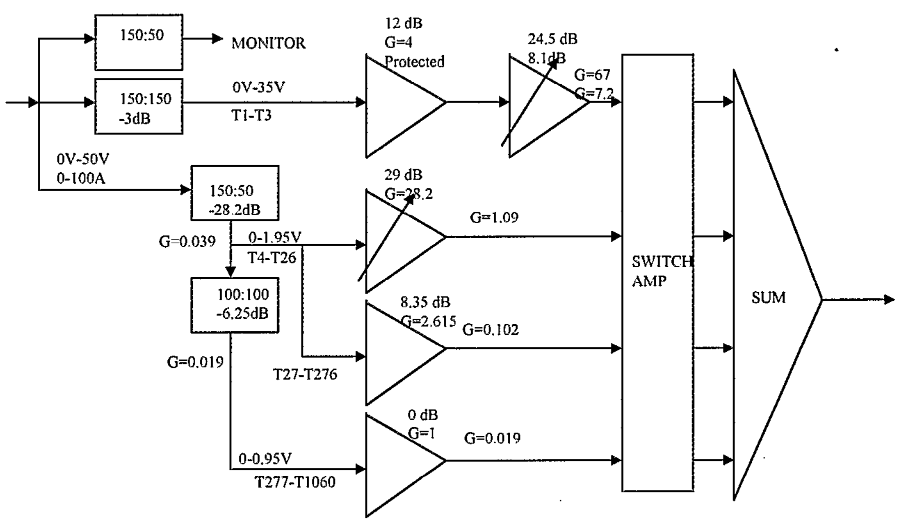


FIGURE 5. Gain Changing configuration

TABLE 1. Gain Switching						
Turn	**Beam Current**	**Est. Input Signal**	**Gain**	**Est. Output Noise**	**Output Voltage**	**Resolution**
1 test	15mA	7.5mV	67	0.65mV	0.5	0.13%
1	38mA	19mV	7.2	0.65mV	0.136	0.47%
4	152mA	76mV	7.2	0.65mV	0.547	0.12%
5	190mA	95mV	1.09	0.485mV	0.103	0.47%
26	988mA	0.494V	1.09	0.485mV	0.538	0.09%
27	1.026A	0.513V	0.102	0.246mV	0.052	0.47%
276	10.488A	5.244V	0.102	0.246mV	0.535	0.046%
277	10.526A	5.263V	0.019	0.246mV	0.1	0.246%
1060	40.28A	20.14V	0.019	0.246mV	0.382	0.064%

Digital control of gain is accomplished via separate DACs to set gain voltage levels on the AD600 variable gain amplifiers, and a gain storage register that establishes which of the gain paths is made active. The gain storage register is updated by a sequence generator on the digital interface section. The gain register information is stored in a FIFO, along with data to tag each data point with a gain path.

ADC Driver Stage

The driver for the ADC has been selected as per the manufacturer's recommendation. An AD8138 wide-band differential ADC drive amplifier is employed to shift the reference to match the reference of the AD6645. The summer feeds the 5 pole Gaussian filter as described earlier. This is buffered by an amplifier and fed to the 17MHz Chebyshev filter.

Digital Processing

The key to the system is the digital processing. This permits us to use transformers that have a very wide bandwidth, and compensate for the droop. The digitized data are transferred to a FIFO for DMA transfer to the PC. A LabVIEW® program processes the data and interfaces the PC with the network.

The DC offset is calculated by averaging points prior to the arrival of the beam. This is subtracted from the data to provide a data set that has been corrected to a zero baseline. The data is then compensated for droop with a digital IIR filter algorithm that cancels the transformer low frequency pole and establishes a new low frequency pole of 1 rad/sec. To cancel the transformer pole, it is necessary to calculate the droop time constant. This is accomplished by providing a calibration pulse and computing the exponential time constant during the transformer recovery time. The sensitivity of resulting compensated droop to errors in this calculation requires a good estimate. An analysis of this indicates a sensitivity of –0.6%/% (error in droop/error in transformer time constant). It is, therefore, necessary to average many data points or calculations.

The sensitivity of the droop to sampling time has a similar sensitivity, +0.6%/% (error in droop/error in sampling time).

The droop compensation formula is:

$$Y(n)=\{1/(2/T+1/\tau_2)\}\{y(n-1)\,(2/T-1/\tau_2) + x(n)\,(2/T+1/\tau_1) + x(n-1)\,(-2/T+1/\tau_1)\}$$

Where: T is the sampling period, $1/\tau_1$ is the transformer lower cut-off radian frequency, and $1/\tau_2$ is the desired new transformer lower cut-off radian frequency.

The data are integrated to determine total charge and filtered for a comfort display. An analysis of integration errors due to insufficient samples indicates a sampling frequency of more than 25MSa/s is required for a 0.1% error in the integral. This analysis was carried out using both a simple sum and a Simpson's rule algorithm. It is interesting to note that the two methods converge at 64MSa/s, and differ only slightly at lower sampling rates. Therefore, there is little advantage to using the more computer intensive Simpson's rule. An example of a simulated beam processed by this software is shown in figure 6.

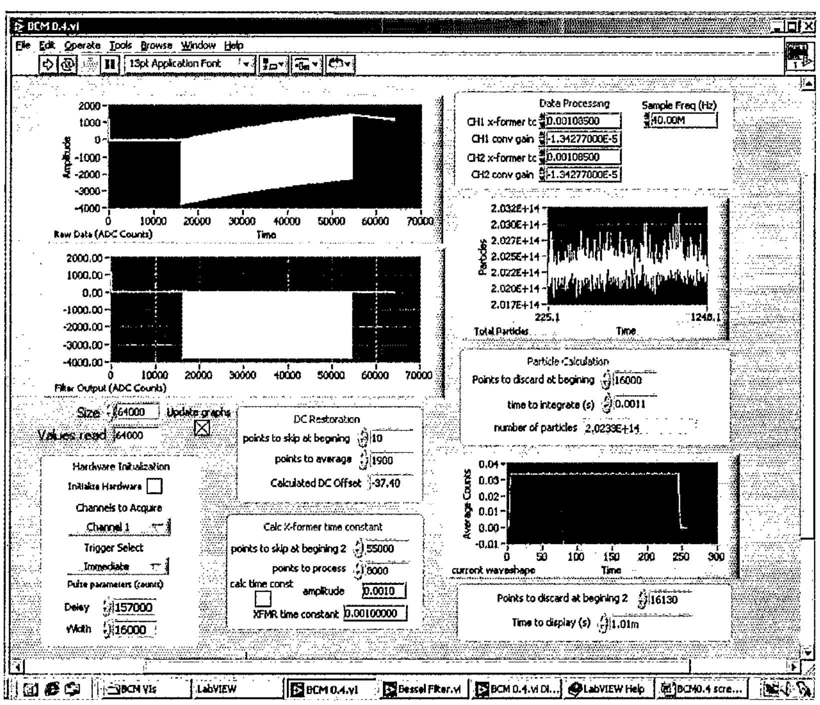


FIGURE 6. Screen dump of a simulated beam processed by the BCM electronics. The upper left graph is raw data showing a simulated 1ms, 645ns pulse with a 945ns period, pulse train. Droop is obvious, and shown compensated in the graph below. The filtered comfort display is shown with 256 points in the lower right, and the integrated charged particle count calculation shown above it.

Testing

A test apparatus was constructed using an 8 inch 50 Ohm coax line. The outer conductor was cut, insulated, and a shroud built to allow the transformer to measure the current in the center conductor. This was found to have a single resonance near 500MHz and provided a good 50 Ohm match for excellent current transient response measurements.

Calibration

An isolated dual, current output, DAC (DAC2902) provides the fundamental source for the calibrator. This is a fast current output device that settles quickly, and can deliver up to 20mA. The calibrator is isolated to avoid ground loops, and is AC terminated to back terminate the calibration winding, while allowing a DC current previously measured by an accurate DMM to flow into the winding. To simulate larger currents for the Ring, an additional current amplifier will be used.

ACKNOWLEDGMENTS

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REFERENCES

1. Kesselman, M. et-al., "SNS Project-Wide Beam Current Monitors", *BIW 2000*, May 8-11, 2000, Cambridge MA.
2. Kesselman, M., et-al. "SNS Project-Wide Beam Current Monitors", *EPAC 2000*, Vienna, Austria, June 26-30, 2000
3. Kesselman, M., "Spallation Neutron Source Beam Current Monitor Electronics", *PAC 2001*, Chicago, Il., June 18-22.